An Optimum Detector for Space-Time Trellis Coded Differential MSK

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ABSTRACT

Space-time (ST) coding using Continuous Phase Modulation (CPM) has spectral advantages relative to linear modulations. In spite of the spectral benefits, Space-Time Trellis Codes (STTC) using the CPM implementation of Minimum Shift Keying (MSK) scheme has inherent inphase and quadrature interference, when the received complex baseband signal is the input into the match-filter to remove the shaped sinusoid pulses. In this paper a novel optimum transmitting and detecting structure for STTC-MSK is proposed. Treating the Alamouti scheme as an outer code, each STTC MSK waveform frame is immediately followed by the orthogonal conjugate waveform frame at the transmit side. At the receiver first orthogonal wave forming is applied, then a new time-variant yet simple trellis structure of the STTC-MSK signals is developed. This STTC-MSK detector is absolutely guaranteed to be I/Q interference-free and still keeps a smaller computation load compared with STTC-QPSK. Simulations are made over quasi-static AWGN fading channel. It is shown that our detector for ST-MSK has solved the I/Q interference problem and has around 2.8 dB gain compared with the Alamouti Scheme and 3.8 dB gain for bit error rate at 5×10^{-3} in a 2 by 1 Multiple Input Single Output system.

KEY WORDS

BER (Bit Error Rate), MIMO (Multiple Input Multiple Output), MSK (Minimum Shift Keying)

INTRODUCTION

Research on wireless communications through multiple input multiple output antennas over none-selective fading channels has received considerable attention [1-3]. Codes design techniques like space-time block codes [5] and space-time trellis codes [6] with linear modulation have been extensively investigated to provide diversity gains and coding gains.

Recently, space-time trellis coded CPM (STTC-CPM) have received a great deal of interest because of the advantage of bandwidth and power efficiency relative to linear modulation. However, owing to the nonlinearity and inherent memory in CPM signal, a direct application of results obtained from the linear modulation to the construction of STTC-CPM is very difficult[7]. In [8], Zhang gave the rank criteria based on [6] for certain STTC-CPM code schemes like STTC-GMSK. A super trellis consisting of the states of space-time codes and the states of partial response CPM was proposed, but the trellis complexity increases exponentially with the length of the CPM memory and STTC states. A newly proposed solution is to apply the delay diversity and update the trellis output every bit time instead of every symbol time for MSK as in [1]. The benefit was that only a two-state trellis was necessary for space-time MSK modulation, but it also introduced inphase and quadrature interference at the detector and caused twice the trellis size expansion at the receiver. The same two state trellis was applied to STTC-GMSK in [9]. With the Phase Amplitude Modulation (PAM) decomposition of GMSK, the simplicity of the modulation trellis structure was inherited, however, the complexity problem and I/Q interference was still unavoidable.

Because MSK belongs to the intersection of CPM and Offset QPSK (OQPSK), it can be viewed as either a case of continuous phase frequency shift keying (CPFSK) with 1REC pulse-shaping function, or a special case of differential Offset QPSK with the half-sinusoidal pulse-shaping. This paper explores a new space-time trellis structure by regarding MSK as a special case OQPSK in the binary space-time trellis coding.

To overcome the difficulties of I/Q interference and correlated noise at the matched filter outputs, Alamouti outer code is added to STTC MSK frame waveform at the transmitter. While at the receiver, the received signal is first stacked according to Alamouti scheme, then a simple time variant space-time trellis is developed to detect the STTC-MSK signal. This trellis detector not only keeps the advantages of simple trellis structure, spectral efficiency, but also removes the inherent I/Q interference and has about 3.8 dB SNR enhancement at bit error rate of 5×10^{-3} and only half the number of the Viterbi branch metrics per bit output and half the number of Viterbi nodes in [1]. Also the computation of the branch metric are all real-based.

Section II models the transmitted signal, Section III describes the derivation of the optimum trellis detection. Section IV gives the simulation of the error performance and section V makes the conclusions.

SYSTEM MODEL

Let us consider a MIMO wireless communication system with M_T transmit antennas and M_R receive antennas. As shown in Figure 1, let $k = 0, 1, 2, \cdots$, be the frame index, the N_F binary bit long frame $\mathbf{b}(k) = [b(kN_F + 1), b(kN_F + 2), \cdots, b(kN_F + N_F)]$ is the input to the space-time encoder to get the coded binary bit streams $\mathbf{c}_i(k) = [c_i(kN_F + 1), c_i(kN_F + 2), \cdots, c_i(kN_F + N_F)], i = 1, \cdots, M_T$, the space-time codeword matrix $\mathbf{C}(k)$ can be denoted by

$$\mathbf{C}(k) = [\mathbf{c}_1(k), \cdots, \mathbf{c}_{M_T}(k)]^T = \begin{bmatrix} c_1(kN_F+1) & \cdots & c_1(kN_F+N_F) \\ c_2(kN_F+1) & \cdots & c_2(kN_F+N_F) \\ \cdots & \cdots & \cdots \\ c_{M_T}(kN_F+1) & \cdots & c_{M_T}(kN_F+N_F) \end{bmatrix}. \quad (1)$$

Then for $n = 1, 2, \cdots, N_F$, $c_i(kN_F + n) \in \{0, 1\}$ is mapped to $v_i(kN_F + n) \in \{-1, 1\}$ using $v_i(kN_F + n) = 1 - 2c_i(kN_F + n), n = 1, \cdots N_F$. Thus, for the differential decoder input matrix codeword $\mathbf{V}(k)$, we define

$$\mathbf{V}(k) = [\mathbf{v}_1(k), \cdots, \mathbf{v}_{M_T}(k)]^T = \begin{bmatrix} v_1(kN_F+1) & \cdots & v_1(kN_F+N_F) \\ v_2(kN_F+1) & \cdots & v_2(kN_F+N_F) \\ \cdots & \cdots & \cdots \\ v_{M_T}(kN_F+1) & \cdots & v_{M_T}(kN_F+N_F) \end{bmatrix}, \quad (2)$$

then modulating symbol matrix $\mathbf{D}(k)$ can be denoted as

$$\mathbf{D}(k) = [\mathbf{d}_1(k), \cdots, \mathbf{d}_{M_T}(k)]^T = \begin{bmatrix} d_1(kN_F+1) & \cdots & d_1(kN_F+N_F) \\ d_2(kN_F+1) & \cdots & d_2(kN_F+N_F) \\ \cdots & \cdots & \cdots \\ d_{M_T}(kN_F+1) & \cdots & d_{M_T}(kN_F+N_F) \end{bmatrix}. \quad (3)$$

The relationship of $v_i(kN_F+n)$ and $d_i(kN_F+n)$ can be established by the differential operation

$$d_i(kN_F + n) = v_i(kN_F + n)v_i(kN_F + n - 1), 2 \leq n \leq N_F \quad (4)$$

The differential decoder between the mapper and MSK is because there is an embedded differential encoder in MSK shown in Figure 3 when it is viewed as OQPSK. Details are in next section.

Each binary data stream $\mathbf{d}_i(k)$ is then used as input toward its MSK modulator. The modulated MSK signal $X(t, \mathbf{d}_i(k))$, $kN_FT_b \leq t < (k + 1)N_FT_b$ can be written as

$$X(t,\mathbf{d}_i(k)) = \sqrt{\frac{E_s}{M_T}}\exp\Big(j\phi(t,\mathbf{d}_i(k))\Big), i = 1,\cdots M_T; \tag{5}$$

and

$$\phi(t,\mathbf{d}_i(k)) = 2\pi\mu\sum_{l=1}^{N_F} d_i(kN_F + l)q(t - kN_F - lT_b) \tag{6}$$

where E_s is the symbol energy, i is the transmit antenna index, $\mu = 1/2$ is the MSK modulation index, $q(t)$ is the correspondent phase shaping function 1REC, T_b is the binary data time interval.

For M_T by M_R MIMO system, the block of the orthogonal waveform may have many various expressions. Let us illustrate with a 2 by 1 or 2 by 2 MIMO system which is actually the Alamouti scheme. The orthogonal wave-forming block inserts a conjugate waveform for every new $N_F * T_b$ long waveform, but different than the Alamouti block coding scheme, the Alamouti transmit scheme here is applied to the modulated waveforms.

During $2kN_FT_b \le t < (2k+1)N_FT_b$ time interval, $X(t,\mathbf{d}_1(k))$ and $X(t,\mathbf{d}_2(k))$ are transmitted over antenna 1 and antenna 2, respectively. During the following $(2k+1)N_FT_b \le t < (2k+2)N_FT_b$ time interval, $-X^*(t-N_FT_b,\mathbf{d}_2(k))$ and $X^*(t-N_FT_b,\mathbf{d}_1(k))$ are transmitted over antenna 1 and antenna 2, respectively.

By doing this, we assume that the channel is quasi-static over $2N_FT_b$ long time interval. For $M_T > 2$, this orthogonal transmit scheme is still possible, but it requires the channel to be quasi-static longer.

OPTIMUM DETECTOR

Since MSK in SISO environment can be thought of as a special case of Offset QPSK with a half-sinusoidal shape and differentially encoded symbols. The optimum MSK detector has a half-sinusoidal matched filter applied to inphase and quadrature modulated signal. The output of the matched filter is then sampled once per bit, or twice per symbol to estimate the symbol transmitted.

However in a MISO or MIMO environment, the complex Gaussian channels and the offset feature of the I/Q waveforms of MSK would unavoidably introduce the interference between inphase and quadrature and also color the white noise, which will severely degrade the performance of the maximum likelihood detector [1].

Orthogonal Wave-Forming

In the proposed transmit scheme, we have the transmitted waveform in Alamouti format, so at the receive side, we can use the orthogonal waveform to get rid of the complex rotation of the channel. for $2kN_FT_b \le t < (2k+1)N_FT_b$, the received signal $y_{2k}(t)$ is

$$y_{2k}(t) = \mathbf{H}\begin{bmatrix} X(t,\mathbf{d}_1(k)) \\ X(t,\mathbf{d}_2(k)) \end{bmatrix} + \mathbf{N}_{2k}(t); \tag{7}$$

where the variance of each element in $\mathbf{N}_{2k}(t)$ is N_o and $\mathbf{H} = [h_1, h_2]$ for a 2 by 1 MIMO system or $H = [h_{11}, h_{12}; h_{21}, h_{22}]$ for a 2 by 2 system. For $(2k+1)N_F T_b \leq t < (2k+2)N_F T_b$, the received signal $y_{2k+1}(t)$ is

$$y_{2k+1}(t) = \mathbf{H}\begin{bmatrix} -X^*(t - N_F T_b, \mathbf{d}_2(k)) \\ X^*(t - N_F T_b, \mathbf{d}_1(k)) \end{bmatrix} + \mathbf{N}_{2k+1}(t). \tag{8}$$

The receiver forms a rearranged waveform vector $\mathbf{Y}(t)$ as

$$\mathbf{Y}(t) = \begin{bmatrix} y_{2k}(t) \\ y^*_{2k+1}(t) \end{bmatrix} = \mathbf{H}_e \begin{bmatrix} X(t,\mathbf{d}_1(k)) \\ X(t,\mathbf{d}_2(k)) \end{bmatrix} + \begin{bmatrix} \mathbf{N}_{2k}(t) \\ \mathbf{N}^*_{2k+1}(t) \end{bmatrix}, \tag{9}$$

where $\mathbf{N}_{2k}(t)$ and $\mathbf{N}_{2k+1}(t)$ are the corresponding complex Gaussian noise respectively. $\mathbf{H}_e$ is the equivalent orthogonal MIMO channel matrix. For example, in a 2 by 1 system, it can be written as

$$\mathbf{H}_e = \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}. \tag{10}$$

The orthogonal property of $\mathbf{H}_e^H \mathbf{H}_e = \|\mathbf{H}\|_F^2 \mathbf{I}$ allows us to multiply the two sides of equation by $\mathbf{H}_e^H$ to obtain

$$\mathbf{Z}(t) = \mathbf{H}_e^H \begin{bmatrix} \mathbf{y}_{2k}(t) \\ \mathbf{y}^*_{2k+1}(t) \end{bmatrix} + \mathbf{H}_e^H \begin{bmatrix} \mathbf{N}_{2k}(t) \\ \mathbf{N}^*_{2k+1}(t) \end{bmatrix}, \tag{11}$$

note that the noise is

$$\begin{bmatrix} \tilde{\mathbf{N}}_{2k}(t) \\ \tilde{\mathbf{N}}^*_{2k+1}(t) \end{bmatrix} = \mathbf{H}_e^H \begin{bmatrix} \mathbf{N}_{2k}(t) \\ \mathbf{N}^*_{2k+1}(t) \end{bmatrix}, \tag{12}$$

which is still white with zero mean and variance $\|H_e\|_F^2 N_o \mathbf{I}$. We have

$$\mathbf{Z}(t) = \|\mathbf{H}\|_F^2 \begin{bmatrix} X(t,\mathbf{d}_1(k)) \\ X(t,\mathbf{d}_2(k)) \end{bmatrix} + \begin{bmatrix} \tilde{\mathbf{N}}_{2k}(t) \\ \tilde{\mathbf{N}}^*_{2k+1}(t) \end{bmatrix}, \tag{13}$$

It is well known that in single input single output (SISO) environment the complex offset QPSK is demodulated by two orthogonal branches of real (in-phase) and imaginary (quadrature) with T_b time offset. In MIMO environment, if we treat space-time trellis coded differential MSK as a space-time trellis coded OQPSK, and Alamouti outer code is applied at the transmitter to counteract the rotation of the channels, then at the receiver, we can form a rearranged vector $\mathbf{Y}(t)$ as in (9) which leads to a decision statistics $\mathbf{Z}(t)$ with a SNR gain of $\|\mathbf{H}\|_F^2$, we can still split the $\mathbf{Z}(t)$ into the branches of real and imaginary and can demodulate the real and imaginary alternately at the receiver.

Offset QPSK Representation of MSK

As shown in Figure 3, the differential decoder before MSK is to compensate the inherent differential encoder embedded in MSK when it is regarded as a differentially coded OQPSK. So the combination of differential decoder and MSK modulator is equivalent to an uncoded Offset QPSK pulse-shaped by sinusoid.

It is well known that MSK has an expression as an offset in-phase and quadrature implementation , i.e, for $kN_F + nT_b \leq t < kN_F + (n+1)T_b$:

$$X(t, d_i(kN_F+n)) = \sum_n \alpha_i(kN_F+n)p(t-kN_FT_b-(2n-1)T_b)-j\sum_n \beta_i(kN_F+n)p(t-kN_FT_b-2nT_b), \tag{14}$$

where

$$p(t) = \begin{cases} \sin\left(\frac{\pi t}{2T_b}\right), & 0 \leq t \leq 2T_b \\ 0, & \text{otherwise.} \end{cases} \tag{15}$$

and

$$\alpha_i(kN_F + n) = (-1)^n a_i(kN_F + 2n - 1); \tag{16}$$

$$\beta_i(kN_F + n) = (-1)^n b_i(kN_F + 2n) \tag{17}$$

$a_i(kN_F + 2n - 1) \in \{-1, 1\}$ and $b_i(kN_F + 2n) \in \{-1, 1\}$ are the odd/even split of a sequence $u_i(kN_F + n) \in \{-1, 1\}$. The sequence $u_i(kN_F + n)$ is related to the sequence $d_i(kN_F + n)$ by

$$u_i(kN_F + n) = d_i(kN_F + n)u_i(kN_F + n - 1), \tag{18}$$

$u_i(kN_F + n)$ is implied in the block of MSK in Figures 3 as an equivalent differentially encoded Offset QPSK. Using the equation 4, we have

$$u_i(kN_F + n) = v_i(kN_F + n), \tag{19}$$

similar to (2), we have the definition of $\mathbf{U}(k)$ with the elements of $u_i(kN_F + n)$, which is omitted for simplicity.

As we can see the transform from differential decoder $\mathbf{V}(k)$ to $\mathbf{D}(k)$ and then differential encoder from $\mathbf{D}(k)$ to $\mathbf{U}(k)$ are is a pair of the transform and its inverse transform. This to say that the input equals output, we can directly use Offset QPSK of $\mathbf{V}(k)$ to denote modulated signal.

The differential decoder at the transmitter could bring a factor of 2 improvement in BER, and allows us to remove the differential decoder at the receiver. This changes the double-error characteristic of MSK modulation to single-error.

With the Offset QPSK representation of MSK, we have $X_{MSK}(t, \mathbf{D}(k))$ written in matrix form

$$X_{MSK}(t, \mathbf{D}(k)) = X_{OQPSK}(t, \mathbf{V}(k)) = \mathbf{V}(k)\mathbf{A}\mathbf{P}(t) \tag{20}$$

where P(t) is N_F by N_F square matrix defined by

$$\mathbf{P}(t) = \mathbf{diag}\{p(t - N_F T_b), \ldots, p(t - N_F T_b - (N_F - 1)T_b)\}. \tag{21}$$

The N_F by N_F square matrix $\mathbf{A}$ is defined by

$$\mathbf{A} = \mathbf{diag}\{1, j, -1, -j, 1, j, \cdots, -1, -j\}, \tag{22}$$

and $\mathbf{V}(k)$ is defined in equation (2) where N_F is assumed to be a multiple of 4.

The detection space-time trellis for $M_T = 2$ is shown in Figure 4. The input to the trellis is the binary information bits $b(kN_F + n)$, but the output for any given n is either the real or the imaginary parts of the complex waveform $X(t, \mathbf{V}(k))$ alternately.

For example, if for the $(kN_F + 4n + 1)$th bit, the previous state is 0 and the input is 1, the output is the real part of the vector $X(t, \mathbf{C}(k))$, i.e. $[0, 1]$ and the space time trellis goes from state 0 to state 1 . Note that the input is updated every T_b time interval, but the output is a 2 by 1 waveform vector with $2 * T_b$ long waveforms for the real or imaginary part of $X(t, \mathbf{C}(k))$. We write the branch output column vectors as row waveform vectors in Figure 5 for notational convenience.

Since the mapper is a linear transform, we can write the MIMO output waveform in matrix form as

$$X_{MSK}(t, \mathbf{D}(k)) = X_{OQPSK}(t, \mathbf{C}(k)) = \exp(j\pi\mathbf{C}(k))\mathbf{A}\mathbf{P}(t). \tag{23}$$

MLSD Detector

Given the representation of MSK as Offset QPSK, the decision statistics in equation (13) be can denoted by

$$\mathbf{Z}(t) = \|\mathbf{H}\|_F^2 \mathbf{V}(k)\mathbf{A}\mathbf{P}(t) + \tilde{\mathbf{N}}(t),, \tag{24}$$

where $\tilde{\mathbf{N}}(t)$ is defined as $[\tilde{\mathbf{N}}_{2k}(t),\ \tilde{\mathbf{N}}^*_{2k+1}(t)]^T$.

The ith ($i \in \{0,1\}$) row of $\mathbf{Z}(t, \mathbf{V}(k))$ can be written as

$$\mathbf{Z}_i(t) = \|\mathbf{H}\|_F^2 X_{OQPSK}(t, \mathbf{v}_i(k)) + \tilde{\mathbf{N}}_i(t)$$

$$= \|\mathbf{H}\|_F^2 \begin{bmatrix} p(t - kN_F T_b) v_i(kN_F + 1) \\ jp(t - kN_F T_b - T_b) v_i(kN_F + 2) \\ -p(t - kN_F T_b - 2T_b) v_i(kN_F + 3) \\ -jp(t - kN_F T_b - 3T_b) v_i(kN_F + 4) \\ p(t - kN_F T_b - 4T_b) v_i(kN_F + 5) \\ jp(t - kN_F T_b - 5T_b) v_i(kN_F + 6) \\ \cdots \\ -jp(t - kN_F T_b - (N_F - 1)T_b) v_i(kN_F + N_F) \end{bmatrix}^T + \tilde{\mathbf{N}}_i(t) \quad (25)$$

where $[\cdot]^T$ is to take the matrix transpose and $\tilde{\mathbf{N}}_i(t)$ is the ith row of $\tilde{\mathbf{N}}(t)$.

The negative signs in the elements can be traced back to (16) and (17). Note that each element of the row vector in the equation (25) at time $kN_F + n$, is the product of $v_i(kN_F + n)$, a delayed pulse-shaping function of $p(t - kN_F T_b - nT_b)$, and a coefficient belonging to $\{1, j, -1, -j\}$.

If we split $\mathbf{Z}(t, \mathbf{V}(k))$ into real and imaginary parts, after sampling the match filter we have

$$\mathbf{Z}(t, \Re\{\mathbf{V}(k)\}) = \|\mathbf{H}\|_F^2 \begin{bmatrix} v_1(kN_F + 1) & v_2(kN_F + 1) \\ -v_1(kN_F + 3) & -v_2(kN_F + 3) \\ v_1(kN_F + 5) & v_2(kN_F + 5) \\ \cdots & \cdots \\ -v_1(kN_F + N_F - 1) & -v_2(kN_F + N_F - 1) \end{bmatrix}^T + \Re\{\tilde{\mathbf{N}}(t)\} \quad (26)$$

for the inphase branch, and

$$\mathbf{Z}(t, \Im\{\mathbf{V}(k)\}) = \|\mathbf{H}\|_F^2 \begin{bmatrix} v_1(kN_F + 2) & v_2(kN_F + 2) \\ -v_1(kN_F + 4) & -v_2(kN_F + 4) \\ v_1(kN_F + 6) & v_2(kN_F + 6) \\ \cdots & \cdots \\ -v_1(kN_F + N_F) & -v_2(kN_F + N_F) \end{bmatrix}^T + \Im\{\tilde{\mathbf{N}}(t)\} \quad (27)$$

for the quadrature branch.

The optimum receiver is the maximum likelihood sequence detector (MLSD). Let $\mathbf{Z}$ be the input signal as defined in equation (11). The optimum estimation of $\hat{\mathrm{b}}$ should satisfy

$$\hat{\mathbf{b}}(k) = \arg\min_{\mathbf{b}} \left\{ \int_{KN_FT_b}^{(K+1)N_FT_B} \|\mathbf{Z}(t) - \|\mathbf{H}\|_F^2 X(t, \mathbf{b}(k))\|^2 dt \right\}. \quad (28)$$

To solve equation (28), Viterbi algorithm is used, where the path metric $M(\cdot)$ changing with time will be

$$M(kN_F + 2n) = \Re\{\mathbf{Z}(t) - \|\mathbf{H}\|_F^2 X(t, \mathbf{b}(k))\} M(kN_F + 2n + 1) = \Im\{\mathbf{Z}(t) - \|\mathbf{H}\|_F^2 X(t, \mathbf{b}(k))\} \quad (29)$$

As mentioned before, Alamouti scheme is applied to counteract the rotation of the channels, which leads to a SNR gain of $\|\mathbf{H}\|_F^2$, at the receiver, $Z(t)$ can be divided into the branches of real and imaginary and can be detected in the real and imaginary alternately at the receiver.

SIMULATIONS

In this section, we present some simulation results to verify the proposed scheme above. The frame lengths are 260 binary bits or 130 QPSK symbols. Each spatial channel is modeled as independent complex AWGN quasi-static channel. Viterbi algorithm is used at the detector.

Figure 6 is the simulation results of 2 by 1 antennas. The performance of Alamouti is from [2]. It shows that our proposed scheme a little steeper slope compared with Alamouti scheme and binary sttc transmit scheme [1] . For bit error rate at 5×10^{-3}, the new trellis detector has 3.8 dB SNR advantage compared with the results in [1] and a 2.8 dB gain with the Alamouti scheme. These prove our purposed detector has no I/Q interference but keeps the diversity order of 2.

The improvement of the performance are from two aspects. First Alamouti scheme is applied here as an outer code of STTC MSK waveform. This makes the complex MIMO channel gain real value and provides the possibility that we can demodulate the space-time coded MSK signal with real part and imaginary part separately. The expectation of real channel gain is $E\|H\|_F^2 = M_T M_R$. Second, by taking the real and imaginary parts separately, the sinusoid wave pulse-shaping filter is fully matched, which neither introduces the I/Q Interference nor color the white complex Gaussian noise.

Now let us compare our trellis with the other two trellises in [6] and [1] respectively. If d is the diversity order, Tarokh's trellis is standard QPSK-oriented, with 4^d branch metrics to update path metrics, while Cavers' trellis is bit-oriented, with 2^{d+2} branch metrics to update two-bit long path metrics. The ratio of $4^d/2^{d+2} = 2^{d-2}$ goes exponentially as the diversity order increases. While our trellis is still bit-oriented but with 2^{d+1} branch metrics to update two-bit long path metrics. This calculation load advantage is fully improved to 2^{d-1} in our algorithm. Also we note that the computation of each metric are all real, which is a fourth of the complex metric.

Finally, to complete a fair comparison, we need to point out that Alamouti transmit scheme is applied to send the conjugate STTC MSK frame signal copy in our proposed scheme. As we can see from the bit error rate simulation that it is well worth the efforts.

CONCLUSION

In summary, this paper provides a new optimum detector trellis structure for STTC-MSK. It is based on the outer code of the Alamouti scheme at the transmitter and the establishment of a new time-variant yet simple trellis structure of the constellations of the STTC-MSK signals at the receiver. This STTC-MSK detector is absolutely guaranteed I/Q interference-free and only has much less the calculation load compared with the known STTC-MSK.

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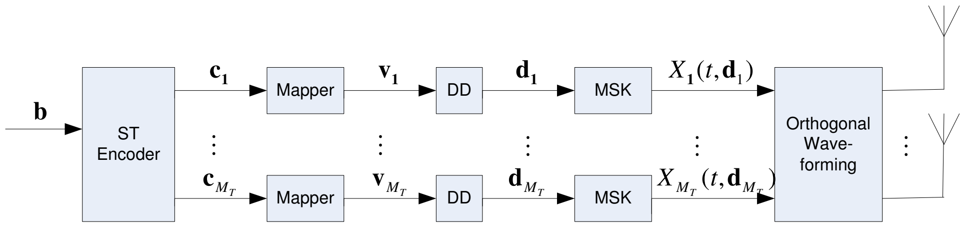

Figure 1: Block diagram of ST-MSK Transmitter

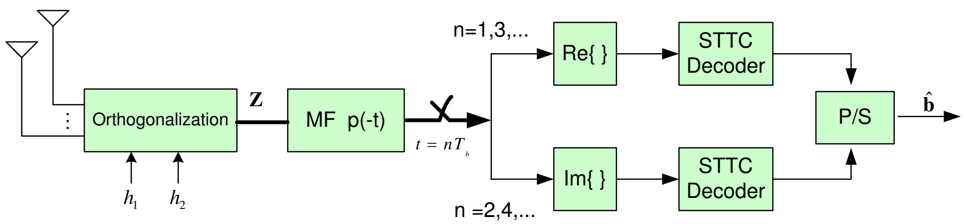

Figure 2: Block diagram of ST-MSK Receiver

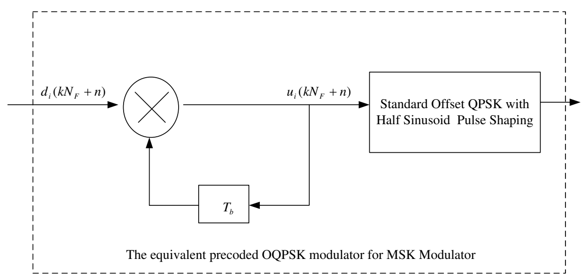

Figure 3: Equivalent Precoded OQPSK for MSK

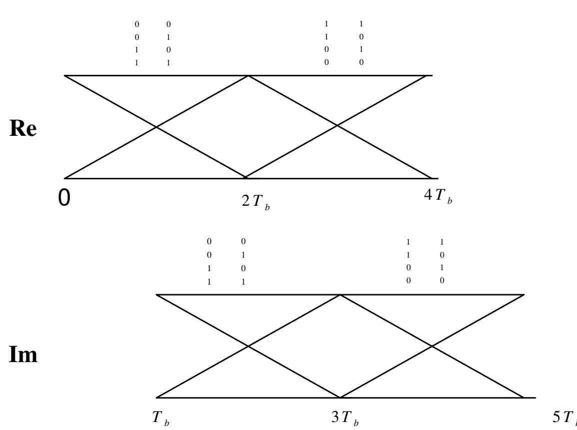

Figure 4: STTC MSK detector trellis for 2 transmit antennas

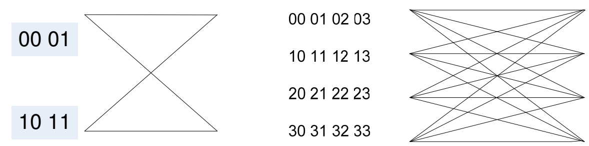

Figure 5: 2 state binary STTC encoder vs. 4 state 4-ary STTC encoder for 2 transmit diversity

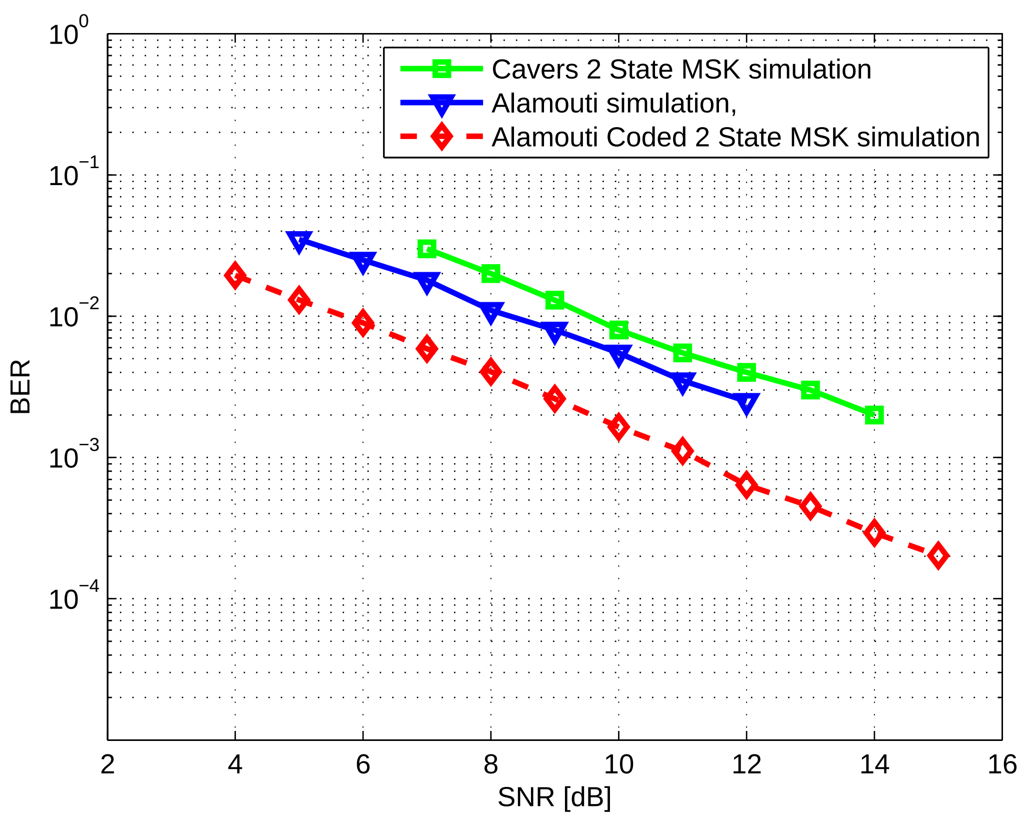

Figure 6: Comparison of BER Performance: Cavers' STTC MSK, Alamouti's scheme and proposed STTC MSK 2 state receivers 2TX 1RX